Tunable Lumped-Element Notch Filter with Constant Bandwidth

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I. Introduction

Interference can drive a receiver's front-end amplifier into compression – distorting, masking, and/or compressing weaker signals of interest – and interferer frequencies can vary with time. Recently, distributed-element absorptive bandstop, or "notch", filters have been demonstrated that can selectively eliminate such interference using lossy (i.e., small and inexpensive) circuit components [1]. These filters maintain excellent characteristics, even when tuned over broad frequency ranges [1]-[3]. To extend this filter technique to lower frequencies, this paper introduces a new lumped-element "absorptive pair" realization that is able to maintain near constant characteristics while tuning over nearly an octave. As shown in Fig. 1, a "first-order" absorptive-pair notch filter [1] consists of a pair of resonators, each coupled to a common transmission line and to each other. Due to their relative simplicity, first-order absorptive filters tend to be the most practical for tunable applications, and they can be cascaded to realize wider stopband bandwidths [1]-[3]. This paper describes such a varactor-tuned lumped-element notch filter – composed of a cascade of three "first-order" stages – with a 26 to 50.375 MHz frequency tuning range, a -30 dB stopband bandwidth of 1.625 MHz, and a -3 dB bandwidth of 5.085 +/- 0.185 MHz.

II. Tunable Notch Filter with Constant Bandwidth

To tune the operating frequency of a filter, reactances of the resonances must be tuned, but care is required to prevent the filter bandwidth from being affected as well. The most practical electronically tunable reactance, an electronically variable capacitance or "varactor", is the tuning element to be used.

A. Lumped-Element Admittance Inverter Couplings

The bandwidth of a parallel-LC resonator notch filter when capacitively tuned to a particular operating frequency is a function of the admittance of both the coupling inverter [4] and the resonator, both of which are functions of frequency. Fig. 2 illustrates that *inductive* Π-type admittance inverter coupling is much better at preserving the bandwidth of a capacitively-tuned single-resonator filter than *capacitive* Π-type admittance inverter coupling. One of the two negative shunt inductors of inverter k_{01} can be absorbed into the shunt resonator inductance, while the other can be absorbed into a shunt inductance of a highpass artificial transmission line [5], such as shown in Figs. 1(b) and 3(a). Note that lumped inductance L_I can also be realized as mutual inductive coupling between shunt inductors L and L_r. To realize an absorptive notch, if k_{01} is inductive, then k_{11} must be capacitive. k_{11} is typically much smaller than k_{01}, and its shunt negative capacitors are small enough compared to the tunable shunt capacitance of the resonators that the absorptive characteristics of the filter of Fig. 1(b) are also preserved over a wide tuning range.

B. Lumped-Element Highpass Transmission Line

The decision to couple resonators to a lumped-element transmission line with inductive admittance inverters mandates the use of a highpass transmission line, as in Fig. 3(a), since it can absorb the $-L_I$ of the inverters. The image-parameter approach of [4]-[6] can be adapted to design a number n of highpass shunt-L/series-C/shunt-L Π-sections to approximate a given physical length d of transmission line at a given frequency f_o. The image impedance Z_i of a single such Π section is [5]:

$$Z_i = \sqrt{L/(2C)} \;/\sqrt{1-(\omega_c/\omega)^2}\,, \qquad (1)$$

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where $\omega_c = 1/\sqrt{2\,L\,C}$ is the highpass cutoff frequency of the Π section in radians/sec, $\omega = 2\pi f$, and f is a frequency in cycles/sec. The image phase shift of n Π sections is:

$$\varphi = 2\,n \sin^{-1}(\omega_c/\omega). \tag{2}$$

Defining a normalized transmission line length S, in terms of a wavelength at operating frequency f_o, corresponding to a transmission line of physical length d by

$$S = f_o\; d\; \sqrt{\varepsilon_r}/c_o\ , \tag{3}$$

where ε_r is the relative permittivity and c_o is the speed of light in a vacuum. The values of L and C for n Π sections approximating a transmission line of characteristic impedance Z_i, S wavelengths long at f_o, can be determined from (1) and (2) to be

$$L = Z_i\,\sqrt{1-\sin^2(\pi S/n)}\ \ /(2\pi f_o \sin(\pi S/n))\ [\mathrm{H}] \tag{4}$$

$$C = \left(4\pi f_o\; Z_i\;\ \sin(\pi S/n)\ \sqrt{1-\sin^2(\pi S/n)}\right)^{-1}\ [\mathrm{F}]. \tag{5}$$

The number of Π sections needed to realize a given length S at frequency f_o with an impedance Zi down to a cutoff frequency of f_c is

$$n = \pi\; S\ /\ \sin^{-1}(f_c/f_o)\ . \tag{6}$$

C. Tuning Element

For convenience, reversed-biased diodes are used as the varactors. An anti-series stacked pair of varactors ideally eliminates varactor-generated second-order distortion and reduces third-order distortion [7]. Stacking multiple anti-series varactor pairs has been recommended to reduce signal voltage across individual varactors and, consequently, signal-induced biasing and distortion [8]. But, such multistack varactor circuits substantially complicate biasing and increase series resistance and parasitics. For filters like those in Fig. 1, a single anti-series pair with larger varactors, or paralleled anti-series varactor pairs as shown in Fig. 4(a), can realize the same reduction in signal voltage across individual varactors as a multistack varactor topology, but with simpler biasing. For a given varactor, increasing the number of stacked varactors reduces the resonator capacitance C_r, requiring an increase in the resonator inductance L_r to preserve the resonant frequency, $\omega_r=(L_r\ C_r)^{-1/2}$. The decrease in C_r and increase in L_r increases the resonator's characteristic impedance, $Z_c\ =\ (L_r/C_r)^{1/2}$, requiring an increase in the coupling-inverter inductance L_I in order to maintain the same effective coupled resonator impedance, $L_{I,new}=(Z_{c,new}/Z_{c,old})^{1/2}\ L_{I,old}$, and preserve the notch bandwidth. The net effect is that the total signal voltage applied to the varactor stack increases by this same amount, $V_{r,new}=(C_{r,old}/C_{r,new})^{1/2}\ V_{r,old}$. Conversely, as the number of paralleled varactor pairs is increased C_r increases, L_r must decrease, Z_c and L_I decrease, and the total voltage applied to the paralleled varactor stack decreases, so that the voltage across individual varactors is actually reduced. Whether a set of N varactors is stacked or paralleled, the signal voltage across an individual varactor in the set, $V_{i,N}$, is a factor of $N^{1/2}$ less than that of a single varactor, $V_{i,1}$, or $V_{i,N}=V_{i,1}/N^{1/2}$. So, to preserve bias simplicity and minimize both parasitics and loss, either a set of paralleled anti-series varactor pairs or a single anti-series varactor pair with larger net capacitance is preferable to multistacked varactors for notch filters comprised of resonators coupled to a common transmission line.

III. The Tunable Third-Order Bandstop Filter

As an example, a lumped-element notch filter with a stopband attenuation of greater than 30dB and a stopband bandwidth of 1.625 MHz was designed to be tunable from 26 to 50.375 MHz. Using (1) - (6), the 50 Ω artificial transmission line phase shift, ϕ, was designed to have a minimum deviation from 90° over the frequency range of interest. The electrical lengths of highpass Π networks of first, second, and third order with minimum deviation from 90° over 24.375 – 52 MHz are plotted as a function of frequency in Fig. 3(b), demonstrating that a second order network as shown in Fig. 3(a), with f_o=32.825 MHz, f_c=12.562 MHz, ϕ=90±34.083°, L=585.275 nH, and C=137.138 pF, represents a good compromise. Next, an MA4ST2600CK-1146T silicon hyperabrupt varactor diode was selected for use in a composite 8-varactor tuning element, as shown in Fig. 4(a), and a series-resistor-inductor-capacitor (series-RLC)

model, shown in Fig. 4(b), was extracted from two-port s-parameter measurements of a 50Ω microstrip line with a shunt-connected reverse-biased varactor diode to ground. Then the circuit of Fig. 1(b), with an unloaded Q of 100 for all inductors, was iteratively-optimized for three operating frequencies: 26, 39, and 50.375 MHz, yielding L_I = 920 nH, C_C = 1 pF, and L_r = 295 nH. Experience with [1] indicated the design should constrain the resonances to be approximately equal at the lowest tuned frequency and constrain one of the two bias voltages to be the highest at the highest tuned frequency. Once the attenuation was more than 50dB at each of the three operating frequencies for some set of bias voltage pairs for the first-order absorptive-pair notch filter, additional offset-tuned first-order stages were cascaded until a 1.625 MHz wide, -30 dB stopband was realized. The resulting cascade of three varactor-tuned absorptive-pair bandstop stages forms the third-order, varactor-tuned lumped-element notch filter shown in Fig. 5(a), in which negative elements have been absorbed into adjacent positive ones. Superimposed plots of the simulated filter characteristics are shown in Fig. 5(b). -30 dB stopband bandwidths are all tuned to 1.625 MHz and resulting absolute 3dB bandwidths are 5.01, 4.91, and 5.27 MHz at 26, 39, and 50.375 MHz.

IV. Conclusion

A varactor-tuned first-order (two-resonator) lumped-element absorptive notch filter with tunable operating frequency and constant absolute bandwidth has been introduced, along with an approach to its design. To reduce signal voltage across individual varactors, it is suggested to add additional pairs of stacked varactors in parallel, rather than in series as advised in [8]. A cascade of three first-order sub-circuits is used to make a third-order filter with useful levels of frequency selectivity and stopband bandwidth and attenuation, with a wide tuning range, and with constant -30dB and -3dB bandwidths. Filters of this type are expected to find use in miniature low-power receivers that must function in dynamic operating environments that may have relatively large narrowband interference.

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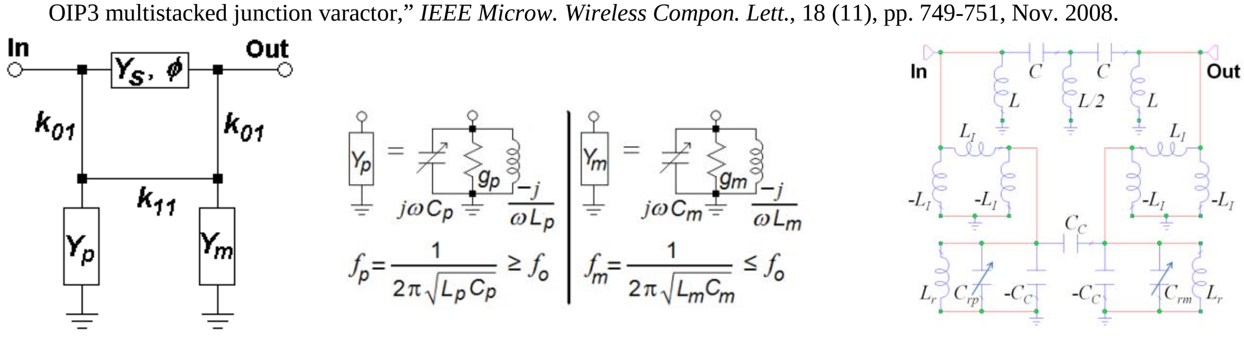


(a) (b)

Fig. 1. (a) An absorptive-pair notch filter with resonator admittances Y_p and Y_m, resonant frequencies f_p and f_m, and tunable capacitances C_p and C_m, each coupled by admittance inverters k_{01} to a common transmission line of admittance Y_S and phase shift ϕ and coupled to each other with inverter k_{11} and (b) the comparable varactor-tuned, lumped-element absorptive-pair notch filter.

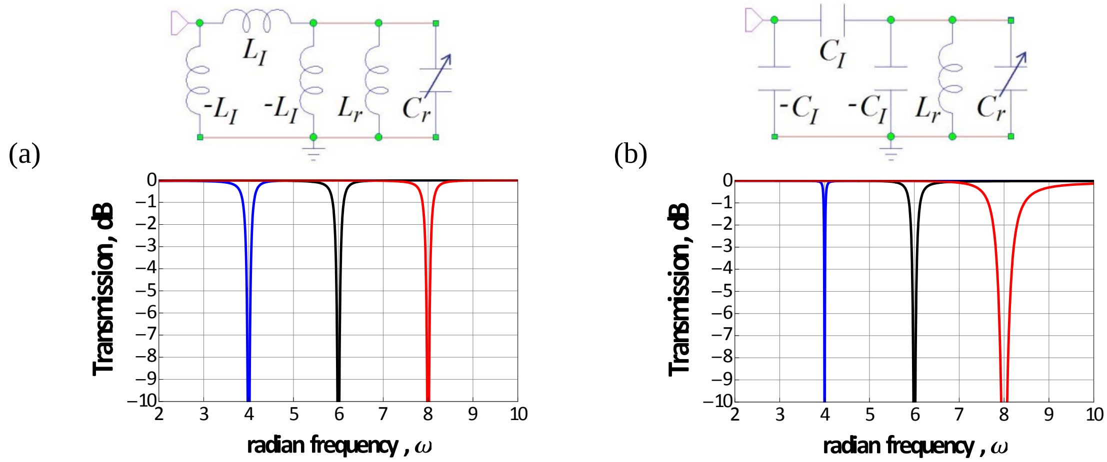


Fig. 2. Schematics and superimposed plots of simulated tuning of capacitively-tuned parallel-LC resonators coupled to a common source and load by (a) inductive and (b) capacitive Π-type admittance inverters (assuming directly-connected 1Ω source and load impedances and component values of L_r=4H, L_I=4H, C_I=0.007F, and $\omega_r = (L_r\,C_r)^{-1/2}$ = {4, 6, 8} radians.

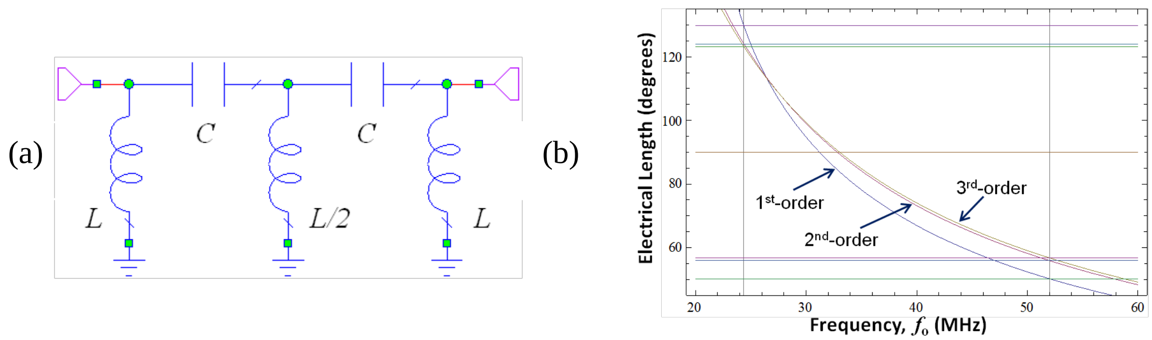
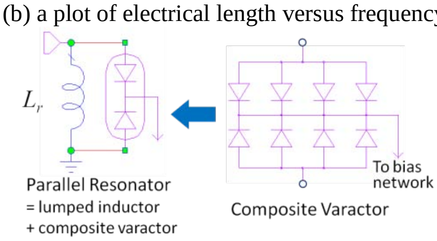
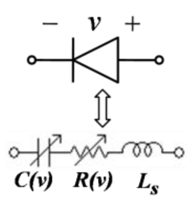


Fig. 3. (a) Equivalent circuit of a lumped-element "artificial" transmission line comprised of a 2nd-order highpass Π network and (b) a plot of electrical length versus frequency for cascaded highpass Π networks of 1st, 2nd, & 3rd order.



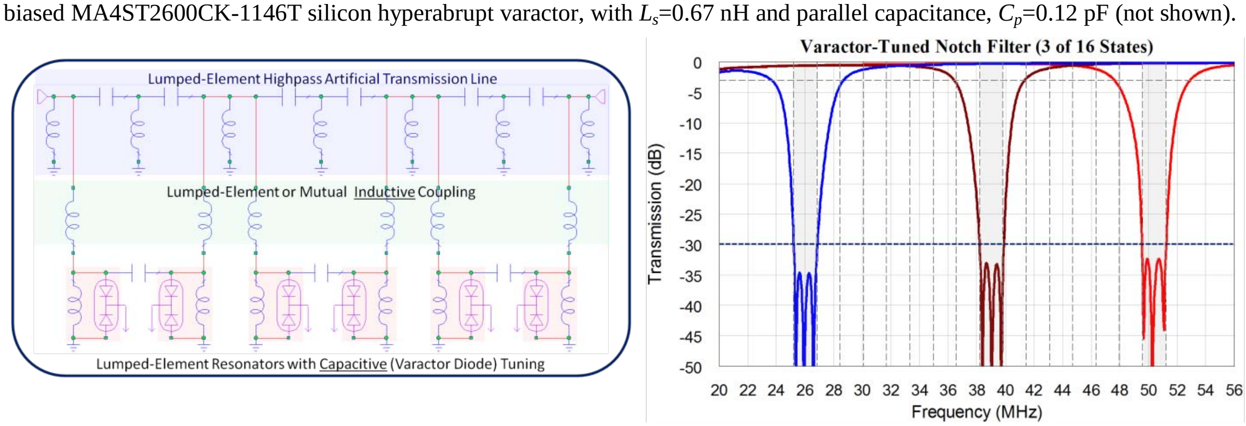


$$R(v)=\begin{cases} 0.519+0.008\,v+0.027\,v^2 & \text{for } 0>v>-1.379 \\ 0.560 & \text{for } -1.379>v>-1.751 \\ 0.396+1.215\,e^{0.577\,v}+\dfrac{0.488}{v} & \text{for } -1.751>v>-5 \end{cases}$$

$$C(v)=6.204+118.901\,e^{-0.0447+0.3554\,v-0.42505\,v^2}$$

(a) (b)

Fig. 4. (a) Capacitively-tuned resonator composed of an inductor and stacked-varactor network and (b) model of the reverse-biased MA4ST2600CK-1146T silicon hyperabrupt varactor, with L_s=0.67 nH and parallel capacitance, C_p=0.12 pF (not shown).



(a) (b)

Fig. 5. (a) Schematic of the tunable, third-order absorptive-pair bandstop filter and (b) superimposed plots of its simulated transmission in three tuned states, demonstrating tunable operating frequency and constant absolute -3dB and -30dB bandwidths.